Communications

On the Accuracy of the Transmission Line Model of the Folded Dipole

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Abstract—**A folded dipole is basically an unbalanced transmission line which radiates as a result of its unbalanced condition. It is known that the folded dipole may he analyzed by considering its current to be composed of two distinct modes, namely, a transmission line mode and an antenna mode. The transmission line model of the folded dipole is investigated with the objectives of determining a) the accuracy of the model and b) the importance of using an equivalent radius in the model. To check the transmission line model, the method of moments is used as a standard for comparison.**

I. INTRODUCTION

A folded dipole is basically an unbalanced transmission line which radiates as a result of its unbalanced condition. It is known [1], [2] that the folded dipole may be analyzed by considering its current to be composed of two distinct modes, namely, a transmission line mode and an antenna mode. A model composed of these two modes will be referred to in this communication as a transmission line model. It will be shown that the input impedance of the folded dipole can be accurately predicted by the transmission line model provided the parallel wires of the folded dipole are sufficiently close together electrically that they function as a normal transmission line. When the wires are not close together electrically, the familiar transmission line equations do not apply. To check the transmission line model, the method of moments is used as a standard for comparison.

II. TRANSMISSION LINE MODEL

To arrive at an equation for the input impedance of the folded dipole, consider Fig. 1, which shows the decomposition of the folded dipole into a transmission line mode current I_T and an antenna mode current I_A. The input impedance Z_T for the transmission line mode will simply be

$$Z_T = jZ_0 \tan k \frac{L}{2} \tag{1}$$

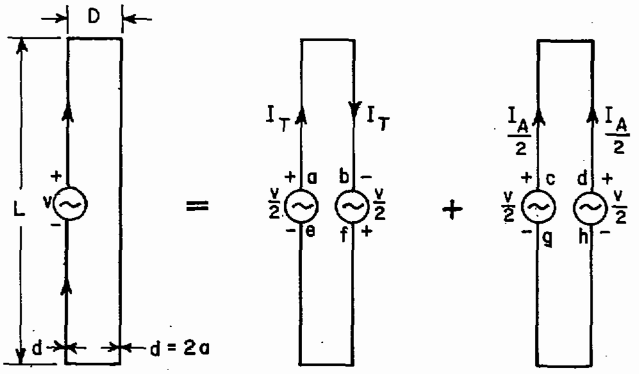

Fig. 1. Decomposition of folded dipole into transmission line mode and antenna mode.

where Z_0 is the characteristic impedance of the line and k is the wavenumber. The transmission line current will be

$$I_T = \frac{V/2}{Z_T} = \frac{V}{2Z_T} \tag{2}$$

since a voltage $V/2$ is applied to the line of length $L/2$ (i.e., the voltage between points a and b is $V/2$). On the other hand, the antenna mode current I_A will be

$$I_A = \frac{V/2}{Z_D} = \frac{V}{2Z_D} \tag{3}$$

where Z_D is the input impedance of a linear dipole of length L. This may be seen from Fig. 1 by connecting the generator terminals of like sign together in the antenna mode drawing (i.e., connect points c and d and points g and h). The result will be a linear dipole made up of two parallel wires extending from the feed point to each (shorted) end of the dipole. The radius of the linear dipole that is used to obtain Z_D should be an equivalent radius determined by the size and spacing of the two parallel conductors [1]. For the cases analyzed here, the equivalent radius a_e is related to the actual wire radius a by

$$\ln a_e = \ln a + \tfrac{1}{2} \ln \frac{D}{a}. \tag{4}$$

To obtain an expression for the folded dipole input impedance, we note that the total current on the "left side" of the folded dipole in Fig. 1 is $I_T + 1/2\, I_A$ and note that the total voltage there is V. Thus

$$Z_{\text{in}} = \frac{V}{I_T + \frac{1}{2} I_A}. \tag{5}$$

Substituting (2) and (3) into (5) yields

$$Z_{\text{in}} = \frac{4Z_T Z_D}{Z_T + 2Z_D}. \tag{6}$$

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Note that, if $L = \lambda/2$, we obtain the well-known result $Z_{\text{in}} = 4Z_D$.

Equation (6) may be used to calculate the impedance of a folded dipole of any length if the input impedance of the corresponding linear dipole is known. In the next section, we will use the transmission line model of (6) to calculate the input impedance of various folded dipoles. The data so generated will be compared against similar data obtained using the piecewise sinusoidal Galerkin moment method for thin wires [3]. The thin wire moment method model used is that suggested by the left side of Fig. 1.

III. RESULTS

Shown in Figs. 2–4 are curves calculated using the transmission line model of the previous section, and also shown is the corresponding moment method model calculation. In each case, the wire radius is 0.0005λ (i.e., $d = 2a = 0.001\lambda$), but the characteristic impedance of the two-wire transmission line used for the folded dipole varies from 300 Ω in Fig. 2 to 600 Ω in Fig. 4.

In Fig. 2, the agreement between the two models is seen to be very good, while in Fig. 4, the agreement is rather poor. The reason for this can be explained in the following way. Since the characteristic impedance of the two-wire line is given by

$$Z_0 = 120 \cosh^{-1} \frac{D}{d} \qquad (7)$$

and therefore depends only on the ratio of D/d or $D/2a$, two closely spaced thin wires can have the same characteristic impedance as two thicker wires having the required larger spacing. Thus in Fig. 5 is presented the input impedance of a folded dipole with $Z_0 = 600\ \Omega$ but with a wire size ten times smaller than in Fig. 4. The agreement in Fig. 5 is seen to be much improved over that in Fig. 4, and this is because the two-wire line is acting as a normal transmission line in Fig. 5 (i.e., (1) applies) but not in Fig. 4, where the wire-to-wire spacing is rather large (i.e., $D = 0.0742\lambda$). We note that in Fig. 2, where agreement is good, $D = 0.00613\lambda$, but in Fig. 3, $D = 0.0213\lambda$ with fair agreement.

The use of the equivalent radius of (4) improves the accuracy of the transmission line model, particularly so for larger wire-to-wire spacings. For example, Fig. 4 also shows results produced by the transmission line model when the equivalent radius is not used. Comparing the two transmission line model curves in Fig. 4 clearly shows the importance of using the equivalent radius in the transmission line model.

IV. CONCLUSION

The results of this investigation show that the transmission line model of the folded dipole will accurately predict the input impedance provided the parallel wires of the folded dipole are electrically close together so that the usual transmission line equations apply [4]. From our observations, a spacing of 0.01λ seems to be a reasonable maximum spacing.

Our results also indicate that it is generally necessary to use the equivalent radius of (4). The necessity of using the equivalent radius depends upon both the wire size and wire spacing (or Z_0). For example, the use of the equivalent radius made very little difference in Fig. 2 but a substantial difference in Fig. 4, where the wire spacing was more than ten times larger.

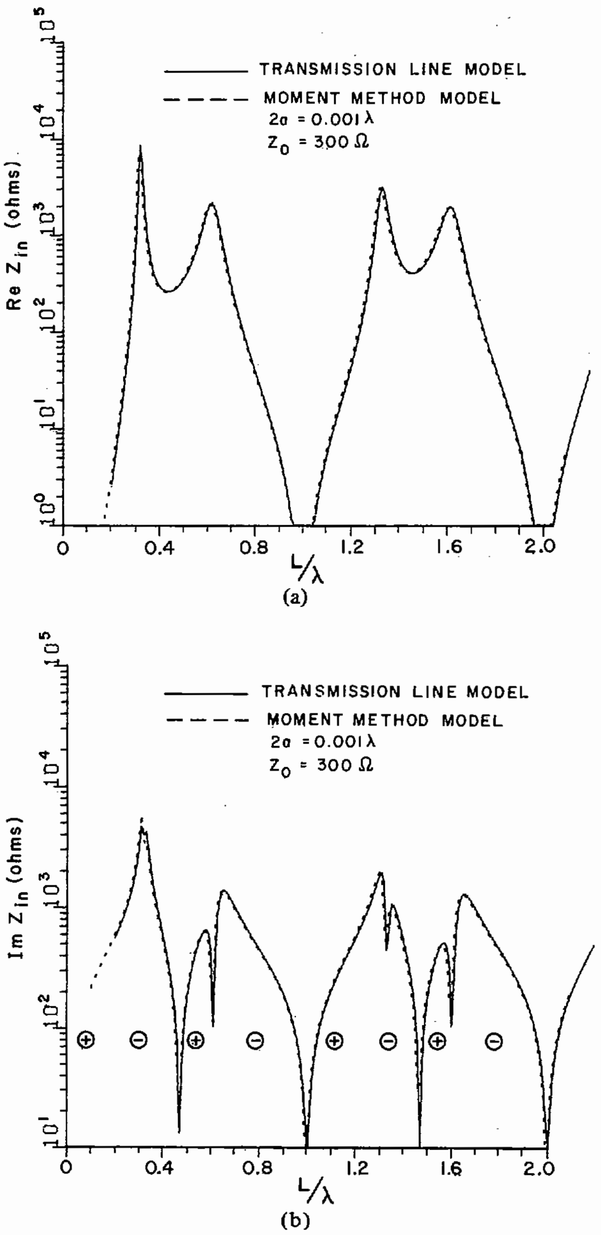
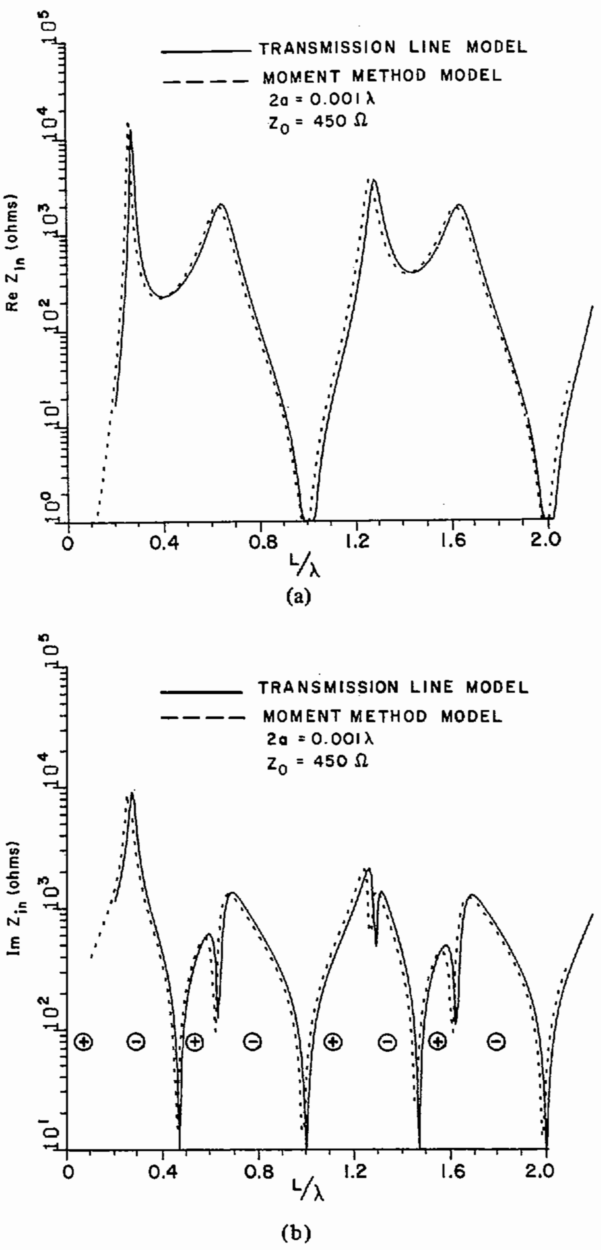


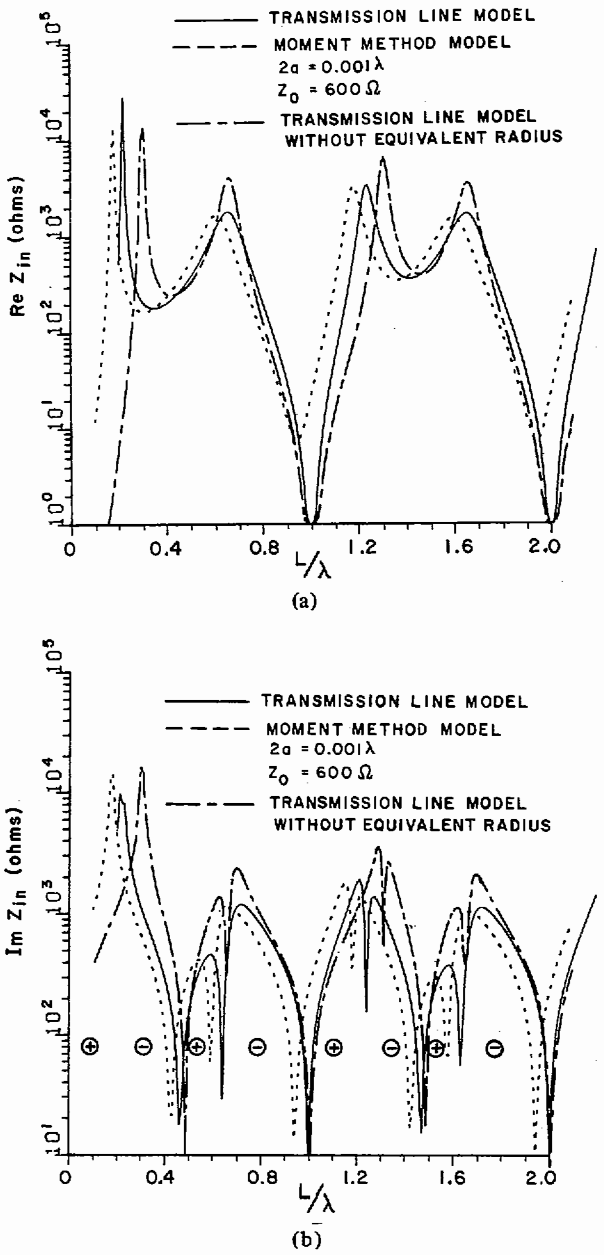
Fig. 2. (a) Comparison of input resistance when $D = 0.00613\lambda$. (b) Comparison of input reactance when $D = 0.00613\lambda$.



(a)

(b)

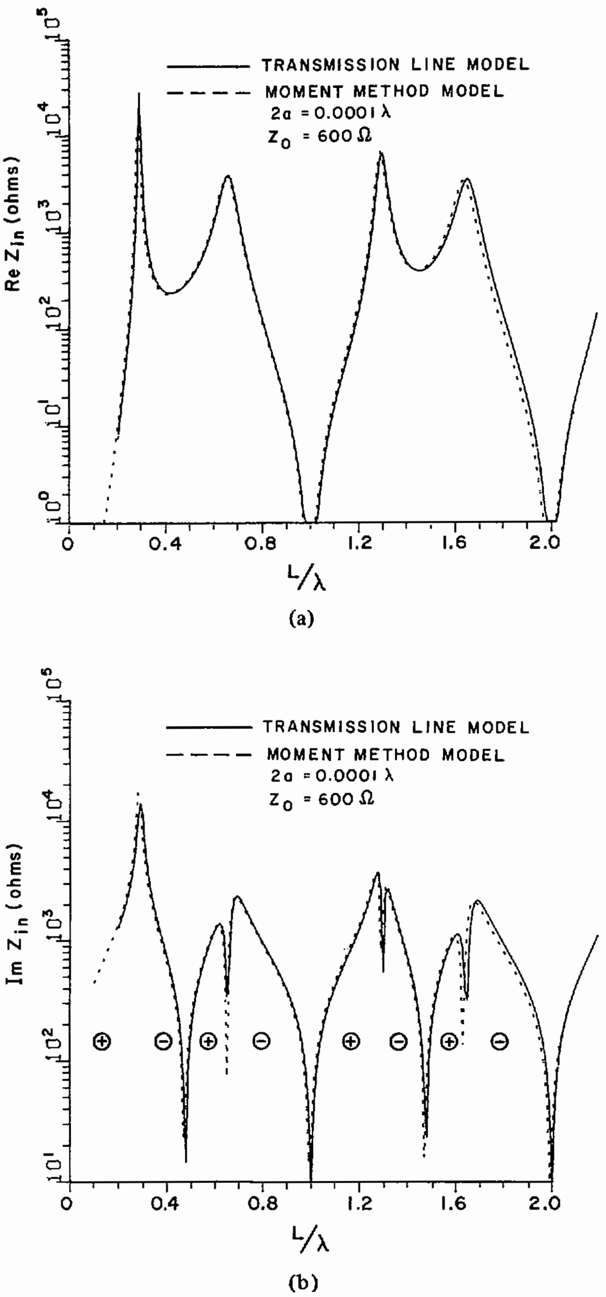
Fig. 3. (a) Comparison of input resistance when $D = 0.0213\lambda$. (b) Comparison of input reactance when $D = 0.0213\lambda$.



(a)

(b)

Fig. 4. (a) Comparison of input resistance when $D = 0.0742\lambda$. (b) Comparison of input reactance when $D = 0.0742\lambda$.



(a)

(b)

Fig. 5. (a) Improvement over Fig. 4(a) for $Z_0 = 600\text{-}\Omega$ case when $D = 0.00742\lambda$. (b) Improvement over Fig. 4(b) for $Z_0 = 600\text{-}\Omega$ case when $D = 0.00742\lambda$.

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Magnetic Polarizability of Some Small Apertures

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Abstract—**Curves are given for the dimensionless magnetic polarizabilities ν_{mx} and ν_{my} of a few characteristic apertures. The relevant integral equations are solved by the moment method. The subareas are triangular, and the basis functions for the triangles touching an edge take the edge singularity into account. Some data are included for a few typical ring-shaped apertures.**

I. INTRODUCTION

In a recent publication [1], data were given concerning the electric and magnetic polarizabilities of small apertures of the type shown in Fig. 1. The accuracy of the polarizabilities was modest (of the order of a few percent) but the data were considered of sufficient interest to warrant publication at the time. The limited accuracy led to some inconsistencies. For example, in the limit $W = l$, diamond and rectangle should yield the same value for the shape factor ν_{mx}, the dimensionless magnetic polarizability. The values obtained in [1] were 0.536 and 0.512, respectively, as compared with an experimental value of 0.518 given by Cohn. The spread, which was due to different patterns of subareas, was brought to the attention of the Editor of this TRANSACTIONS by Prof. W. A. Davis of Virginia Polytechnic Institute. In the meantime, much better algorithms were developed at this Laboratory which yield, e.g., a common value of 0.519 for square and diamond. It was decided by all concerned that publication of improved curves would be in the interest of the scientific community. These curves are found in Figs. 4 and 5. They are only concerned with the magnetic polarizability. Improved data concerning the electric polarizability will be published later. We have also added a few data relative to the ring-shaped apertures shown in Fig. 2. The problem is of interest for the evaluation of microwave power leaking through cracks around doors and the penetration of strong electromagnetic pulses through imperfectly shielded windows.

II. NUMERICAL METHOD

The problem in hand is to determine a dimensionless polarization dyad $\bar{\bar{\nu}}_m$, in terms of which the magnetic moment of the aperture can be written as

$$\bar{p}_m = -S^{3/2}\bar{\bar{\nu}}_m \cdot \bar{H}_g. \tag{1}$$

In this formula, S is the area of the aperture. It is assumed that the aperture is illuminated from one side only (say side 1). For such a case, $\bar{H}_g$ is the (short circuit) field on side 1 of the aperture, the latter being metallized. The fields on the other side (side 2) are produced by the free space moments $\bar{p}_m$ and $\bar{p}_e$ (the electric dipole moment). Generalization to screens separating two different media, in each of which volume sources may exist, is immediate [1].

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